4.8 A 3.2-to-7.3GHz Quadrature Oscillator with Magnetic Tuning

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LC voltage-controlled oscillators are largely adopted, in narrow-band wireless applications, to generate the internal reference of transceiver ICs, due to their remarkable spectral purity at relatively low power levels. In the framework of wideband applications, e.g., ultra-wideband and multi-mode multi-standard systems, duplicate oscillators is the most natural choice to cover the entire bandwidth, while saving power, because ring-based solutions display poor phase noise [1,2]. In this paper, a quadrature oscillator is introduced that employs a frequency tuning method, based on the control of the magnetic field of a transformer, allowing continuous frequency variation over a very wide range. Prototypes, integrated in a 65nm CMOS process, show 3.2 to 7.3GHz frequency tuning range, a phase-noise figure of merit of 176.5dB at 3.2GHz, 170.5dB at 6.4GHz, and 164dB at 7GHz, and a phase error from quadrature <1.5°.

Referring to Fig. 4.8.1, the key idea is that the resonance frequency (f_0) of a transformer-capacitor network can be tuned varying the current through the transformer secondary winding. The voltage across the primary winding depends on both i_1 and i_2, via the mutual inductance (M) and the self inductance (L_1), respectively. So does the current at resonance. Assuming $\alpha=i_2/i_1$ is a real number, f_0 is given by:

$$f_0 = \frac{1}{2\pi\sqrt{(L_1+\alpha M)C}}$$

Depending on the sign of α, the resonance frequency can be tuned either in an upper or in a lower band around $1/(2\pi\sqrt{L_1C})$. Interestingly, the upper band extends theoretically to infinity, when $i_2/i_1 = L_1/M$.

When loading a negative resistance pair to realize an oscillator, the transformer-capacitor network allows frequency tuning, provided the currents in the two windings are exactly at the same frequency and aligned in phase or anti-phase. Two identical oscillators with quadrature outputs are suited to implement this function. Referring to the coupled oscillators of Fig. 4.8.1, assuming they run in quadrature, the phases of the two currents in each transformer are forced to align. For example, referring to Osc A, i_{2B} is in phase with V_A and i_{1B} is in quadrature with V_B.

The proposed magnetic coupling of the two oscillators, directly connected in one direction and cross connected in the other, guarantees quadrature outputs. If $\alpha > 0$, mutual and self voltage components are in phase and if $\alpha < 0$ they are anti-phase.

Selection of the wider up-frequency tuning band is achieved, purposely introducing a negative phase shifter, e.g., an RC filter, in both connecting paths. Mutual magnetic coupling induces a resistive component. By inspection, the resistance is positive in the lower band, while it is negative in the upper one and thus reinforcing the oscillation mode.

The transistor-level schematic of the differential quadrature oscillator is shown in Fig. 4.8.2. Transistors $M_{1,2}$ implement the negative resistance, while $M_{3,4}$ realize the coupling transconductors. Assuming hard switching during oscillation, the pairs produce square-wave currents. The equivalent model of Fig. 4.8.2 captures the large signal behavior of each oscillator in the coupled structure. By inspection of the circuit, where for the sake of simplicity the transformer quality factor is limited by series losses only ($Q_T=\omega L_1/R_{s1}$), the 0-peak oscillation amplitude is derived as $V=2Q_TI_{core}/(\pi\omega_0C)=2L_1I_{core}/(\pi R_{s1}C)$. In the entire tuning band, the output amplitude does not change with frequency and depends on design parameters only.

The transformer, detailed in Fig. 4.8.3, is designed with compromise between quality factor, self-resonance, and coupling factor ($k=M/\sqrt{L_1L_2}$). To maximize M with respect to L_1, L_2 is chosen higher than L_1, roughly twice. A differential structure, with central tap, is used. The two spirals with patterned ground shield are interleaved. The primary has 3 turns and the secondary has 5. The quality factor of the primary spiral is a key design parameter. Therefore, for the primary winding, the upper three metal levels are used and the trace has a relatively large width. On the contrary, the quality factor of the secondary has minor impact on the performance. To minimize fringing and substrate capacitances, a smaller width and only the two upper levels are used. The quality factor of L_1 and the coupling factor, simulated with ADS Momentum, are also shown in the figure. The peak Q is 11, k is higher than 0.8 in the entire frequency range and the self-resonance is 12.6GHz.

The minimum oscillation frequency is chosen as 3.2GHz, determining a maximum capacitance value of 1.4pF. The maximum oscillation frequency is limited by the transformer loaded self resonance. To further extend the tuning band a MOS variable capacitor is used. The noise current of M_6, directly modulating the tuning current, is minimized choosing a non minimum length, large area device (W/L = 800µm/1µm). The RC network, used as a phase shifter to select the tuning band, is implemented by means of the MOS parasitic gate resistor and gate capacitor. Few degrees of phase shift guarantee selection.

The quadrature oscillator, fabricated in a 65nm CMOS process, has been bonded on a printed circuit board for characterization. Figure 4.8.4 shows the chip micrograph. The die area including bond pads is $1mm^2$, with the core cell occupying $0.2mm^2$. Parts of a single-ended slightly modified version of the transformer have also been characterized. Measurements and simulations of Q, k, and self-resonance differ by less than 5%. The quadrature oscillator is supplied at 1.2V during measurements. The oscillation frequency is plotted in Fig. 4.8.5 versus I_{tune}, for I_{core} = 2.5mA, when the tank capacitance is maximum and when it is minimum. Reducing I_{core} to 1.25mA allows exploring the oscillation frequency up to the limit set by transformer loaded self resonance. The minimum oscillation frequency is 3.2GHz while the maximum is 7.3GHz, which are in very good agreement with simulations. The measured output power when I_{core} = 2.5mA, de-embedding the attenuation due to the output buffer and experimental setup, is 0dBm and is roughly constant in the entire band. The measured phase noise is reported in Fig. 4.8.6, at 3.2GHz, 6.4GHz, and 7GHz for I_{core} = 2.5mA, and at 7GHz for I_{core} = 1.25mA. At higher frequency, the noise contribution of the tuning circuit increases. The phase-noise figure of merit (FOM) is determined according to the expression introduced in [3]. When calculated at 10MHz offset from 3.2GHz, 6.4GHz and 7GHz, it is given by 176.5dB, 170.5dB, and 164dB, respectively. To determine quadrature accuracy, two purposely designed downconversion mixers are driven by the quadrature oscillator. An externally provided RF signal is downconverted in quadrature at 1MHz and the phase difference is detected. In the entire band, the phase deviation from quadrature is lower than 1.5°. Measurement results are summarized in the table shown in Fig. 4.8.4.

Acknowledgment:
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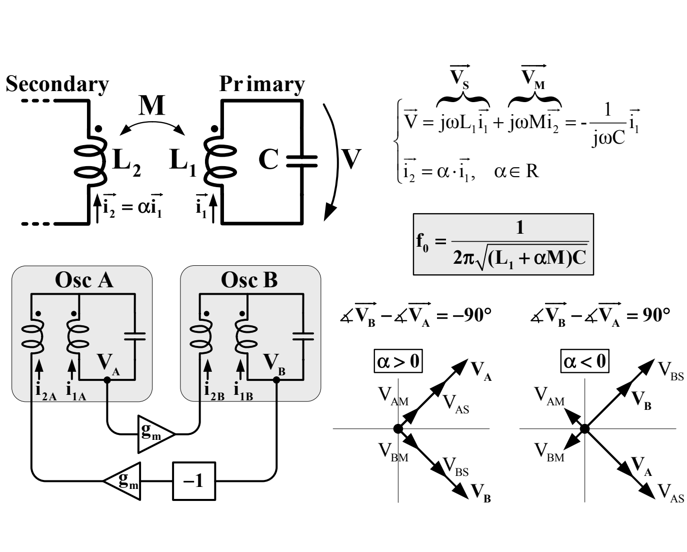

Figure 4.8.1: Magnetic tuning principle.

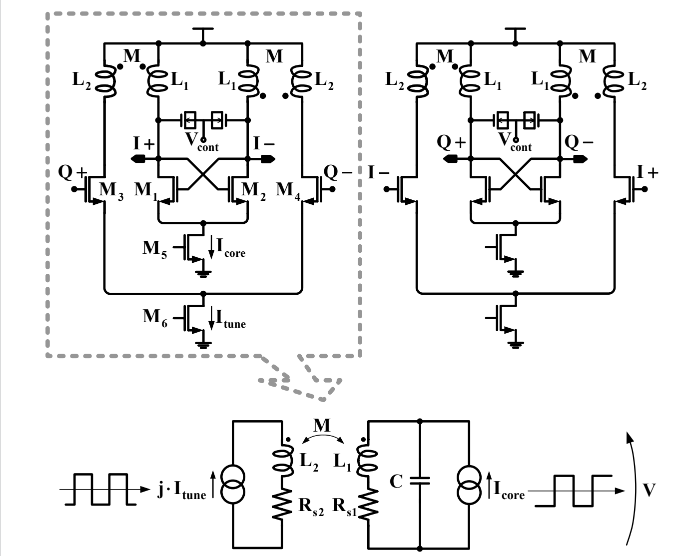

Figure 4.8.2: Quadratureoscillator schematic and large-signal equivalent model.

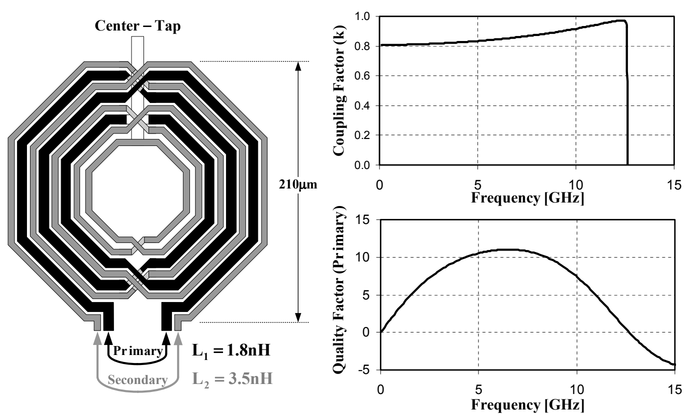

Figure 4.8.3: Transformer layout and Simulated coupling (k) and quality (Q) factors.

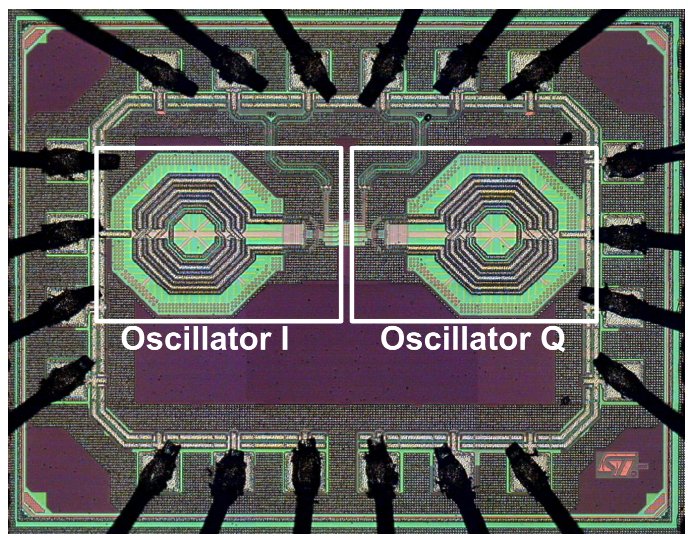

Figure 4.8.4: Chip micrograph.

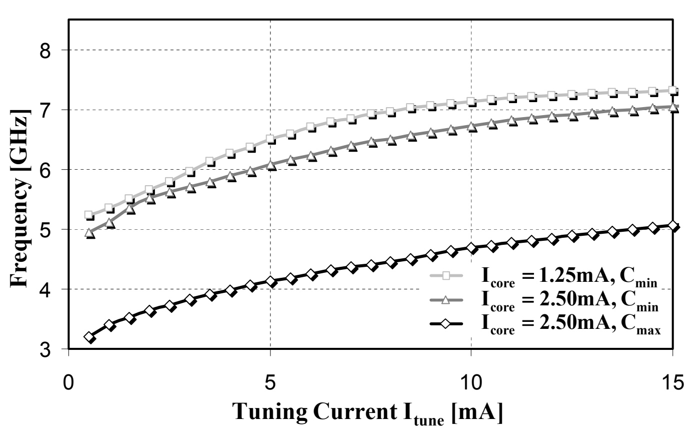

Figure 4.8.5: Oscillation frequency versus tuning current.

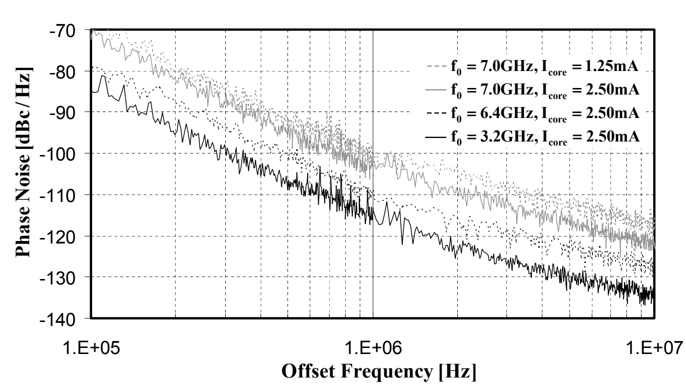

Figure 4.8.6: Phase noise versus frequency offset.

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Frequency Range [GHz]	3.2 to 7.3
Phase Noise FOM @ 3.2GHz [dB]	176.5
Phase Noise FOM @ 7.0GHz [dB]	164
Phase error from quadrature [°]	< 1.5
Voltage Supply [V]	1.2
Current Consumption [mA]	6 to 35
Die area [mm^2]	1
Active area [mm^2]	0.2
Technology	65nm CMOS

Figure 4.8.7: Measurement summary.

